Sniffing EMI in the Near Field

FEATURE ARTICLE

David Prutchi

Noncompliance with FCC and EMC regulations affects everyone! So, David shows designers how to construct probes that identify and isolate sources of EMI that inevitably make it into a circuit.

The technologies involved in modern circuit design have considerably blurred the boundaries between digital and analog worlds. Suddenly, 100-MHz clocks became commonplace in high-performance digital circuits. Now it's necessary to consider every connection between components as an RF-transmission line.

As the need for higher performance pushes designers toward high-speed technology, the market demands more compact, lighter, and less power-hungry devices. With smaller size, analog effects again enter into consideration.

As components and conductors come into close proximity, coupling between circuit sections becomes a problem. Obviously, self-interference must be eliminated to make the product workable, but this still doesn't make the product market worthy.

To ensure that devices do not interfere with each other, strict regulations concerning electromagnetic compatibility are now enforced worldwide.

In the U.S., the FCC regulates the testing and certification of all electronic devices which generate or use clock rates above 9 kHz [1]. In principle, the FCC's charter protects communications from unwanted electromagnetic interference (EMI).

In the European Union, an Electromagnetic Compatibility (EMC) Directive prohibits undue interference to radio and telecommunications equipment. Equipment must possess sufficient immunity to operate as intended in the presence of interference [2].

Designers must follow these requirements. Failure to comply with EMI and EMC regulations has a serious impact for everyone associated with the product—designers, manufacturers, marketing, distribution, and even customers.

Noncompliance can halt manufacturing and distribution, incur fines, and cause the public posting of notices of noncompliance to warn potential customers and other agencies [3].

Assuring compliance involves an extensive series of tests. The EMI and EMC standards clearly define the construction of test sites, as well as the necessary test procedures. Even a fairly Spartan facility capable of conducting these tests can cost over $100,000 just to set up. Most companies, therefore, hire an outside test lab at $1000–2000 per day to conduct testing.

Considering how fast charges accumulate during testing, it's obviously not smart to simply hire a test lab and wait for the results. Rather, designers should familiarize themselves with the relevant EMI and EMC standards and consider the compliance requirements at every stage in the design process.

In *INK* 61, Jeff Bachiochi presented the basics of designing digital circuits for compliance, as well as some methods for troubleshooting circuits to reduce potential problems at the time of testing for compliance.

In this article, I delve deeper into the theory of how digital circuits produce EMI. I also describe some low-cost tools and methods so you can identify and isolate the EMI sources that inevitably make it into a circuit.

RADIATED EMISSIONS FROM DIGITAL CIRCUITS

Digital circuits constantly switch the state of lines between high- and low-voltage levels to represent binary states.

Figure 1a shows that the resulting time-domain waveform on any single line of a digital circuit can be idealized as a train of trapezoidal pulses of amplitude A (either current [I] or voltage [V]), risetime (t_r), falltime (t_f) at 10–90% amplitude, pulse width (τ) at 50% amplitude, and period (T).

The Fourier envelope of all frequency-domain components generated by such a periodic pulse train can be

approximated by the nomogram in Figure 1b. The frequency spectrum is mainly a series of discrete sinewave harmonics starting at the fundamental frequency $f_0 = 1/T$ and continuing for all integer multiples of f_0.

The nomogram identifies two frequencies of interest. At f_1, the locus of the maximum amplitudes rolls off with a $1/f$ slope. At f_2, the locus rolls off at a more abrupt rate of $1/f^2$. These frequencies are located at:

$$f_1 = \frac{1}{\pi\tau} = \frac{0.32}{\tau}$$

and

$$f_2 = \frac{1}{\pi t} = \frac{0.32}{t}$$

where t is the faster of (t_r, t_f).

The envelope of harmonic amplitude (in amps or volts) then simplifies to:

$$|\,V \text{ or } I\,| = \frac{2A(\tau + t)}{T}$$

where f is less than f_1,

$$|\,V \text{ or } I\,| = \frac{0.64\,A}{Tf} = -20\text{ dB / decade roll off}$$

where f_1 is less than or equal to f, and f is less than f_2, and

$$|\,V \text{ or } I\,| = \frac{0.2\,A}{Tf^2} = -40\text{ dB / decade roll off}$$

where f_2 is less than or equal to f.

For nonperiodic trains, the nomogram must be modified to account for the broadband nature of the source. To do so, define a nomogram of the spectral-density envelope of the signal for a unity bandwidth of 1 MHz by:

$$|\,V \text{ or } I\,|\left[\frac{dBV}{MHz} \text{ or } \frac{dBA}{MHz}\right] = 6 + 20\log(A\tau)$$

where f is less than f_1,

$$|\,V \text{ or } I\,|\left[\frac{dBV}{MHz} \text{ or } \frac{dBA}{MHz}\right] = 20\log(A) - 4 - 20\log(f\,[MHz])$$

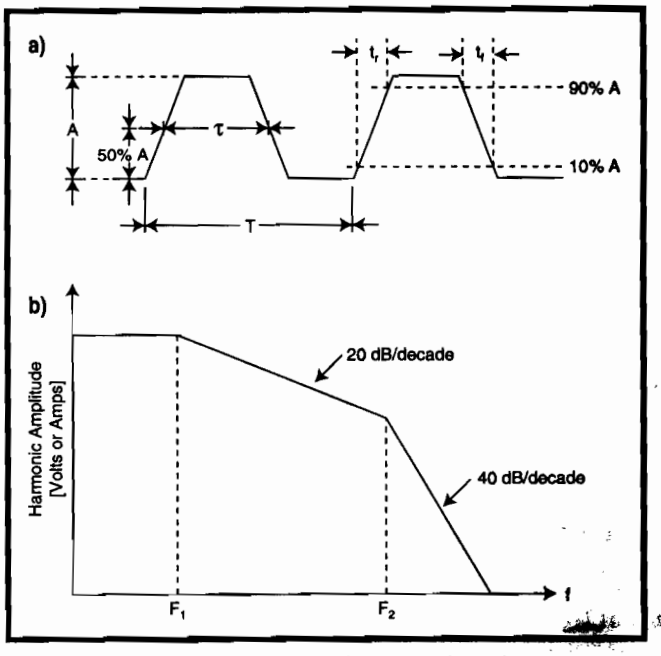


Figure 1—*A pulse train with the characteristics shown in (a) produces a spectrum with an envelope that can be approximated by the nomogram of (b).*

where f_1 is less than or equal to f, and f is less than f_2, and

$$|\,V \text{ or } I\,|\left[\frac{dBV}{MHz} \text{ or } \frac{dBA}{MHz}\right] = 20\log\left(\frac{A}{t\,[\mu s]}\right) - 14 - 40\log(f\,[MHz])$$

where f_2 is less than or equal to f.

Depending on its internal impedance, a circuit carrying a pulse train creates a field in its vicinity which is principally electric or magnetic. At a greater distance from the source, the field becomes electromagnetic, regardless of the source impedance.

If there is a conduction or radiation coupling mechanism, some or all of the frequency components in the digital pulse train's spectrum are absorbed by a "victim" receiver circuit.

To illustrate the extent of the problem, imagine a microcomputer motherboard consisting of a CPU, glue logic, and memory ICs in an unshielded plastic case.

Assume that a number of these ICs toggle states synchronously at a frequency of 100 MHz. Also, assume the total power switched by the circuit at any instant during a synchronous transition is approximately 10 W. Since a real circuit's efficiency is not 100%, a small fraction of this 10 W does no useful work, is not dissipated as heat by the ICs and wiring, but radiates into space.

The power radiated is 10 μW since a reasonable fraction value of 10^{-6} equals the total switched power at the fundamental frequency. If an FM radio is placed 5 m from the motherboard, the field strength E produced by the 10 μW at this distance is approximated by:

$$E = \sqrt{\frac{30\,RadiatedPower[W]}{Distance[m]}} = \sqrt{\frac{30\times 10\times 10^{-6}}{5}} = 3.46\left[\frac{mV}{m}\right] = 70.79\left[\frac{dB\mu V}{m}\right]$$

Technology	Minimum Voltage Swing [V]	Minimum Transition Time t [ns]	Typical Bit Pulse Width τ [ns]	Equivalent Bandwidth [MHz]	Single-load Input Capacitance [pF]	Output Source Impedance (Low/High)
5-V CMOS	5	70	500	4.5	5	300/300
12-V CMOS	12	25	250	12	5	300/300
HCMOS	5	3.5	50	92	4	160/160
TTL	3	8	50	40	5	30/150
TTL-S	3	2.5	30	125	4	15/50
TTL-LS	3	5	50	65	5.5	30/160
TTL-FAST	3	2.5	25	125	4.5	15/40
ECL	0.8	2	20	160	3	7/7
GaAs	1	0.1	2	3200	1	N/A

Table 1—*The most popular logic families have very different timing and driving parameters, resulting in radiated emissions spectra with different characteristics.*

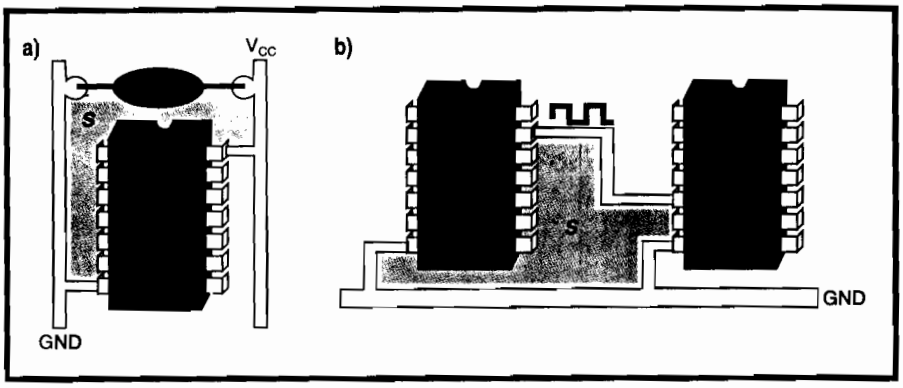


Figure 4—*Simple differential-mode radiating-circuit configurations are created when an AC current flows on a current path that forms a loop enclosing a certain area S. a) Transient power demands of an IC are supplied by a decoupling capacitor, causing brief strong currents that circulate on a loop formed by the supply-bus PCB tracks. b) Fast digital signals driving low-impedance inputs form EMI-radiating loops when current returns through distant ground paths.*

If the field is generated by a high-current, low-voltage circuit, the field is mostly magnetic in nature. If, on the other hand, the field is produced by an element placed at high voltage with little or no current, the field is mostly electric in nature. This is the domain of the near field, while the plane wave is in the domain of the far field.

The ideal generator for a magnetic field (*H-field*) is thus a circular loop of area $S[m^2]$ carrying an AC current I of wavelength λ.

Note that, although a static field is generated by a DC current and can be calculated with the following method, static H-fields do not cause radiated emissions. They are thus disregarded for EMI purposes.

If the loop size is smaller than the observation distance D, the magnitudes of the $\vec{E}$ and $\vec{H}$ vectors are found using the solutions derived from Maxwell's equations.

In the near field, the simplified values for these magnitudes are:

$$H\left[\frac{A}{m}\right] = \frac{IS}{4\pi D^3}$$

and

$$E\left[\frac{V}{m}\right] = \frac{Z_0 IS}{2\lambda D^2}$$

where Z_0 equals the impedance of free space, 120π, or 377 Ω.

Inspecting these equations, we find that in the near field, H is independent of λ and decreases drastically with the inverse of the distance cubed. At the same time, the electric field increases as frequency increases, and it falls off with the inverse of the square of distance.

The wave impedance may be defined as the division of E by H because:

$$Z_{wave}[\Omega] = \frac{E\left[\frac{V}{m}\right]}{H\left[\frac{A}{m}\right]}$$

thus, in the near field,

$$Z_{wave} = \frac{Z_0 2\pi D}{\lambda} \qquad (1)$$

where

$$D < \frac{48}{F[MHz]}$$

In the far field, on the other hand, both E- and H-fields decrease as the inverse of the observation distance as described by:

$$H\left[\frac{A}{m}\right] = \frac{\pi IS}{\lambda^2 D}$$

$$E\left[\frac{A}{m}\right] = \frac{Z_0 \pi IS}{\lambda^2 D}$$

which maintains a constant impedance equal to Z_0. You can therefore directly calculate the radiated power density in W/m^2 by multiplying E an H.

E and H, and thus power, increase with the square of frequency. Limitir the bandwidth of radiated signals by pulse train is therefore of utmost importance in controlling EMI.

The region dividing the near field from the far field is called the *transition region* (i.e., at $D = 48/f$ *[MHz]*). Abrupt transitions occur in near-fiel characteristics until a smooth blendi leads to far-field characteristics.

Electromagnetic fields are also cr ated by passing an alternating curren through a straight-wire dipole, just a with a radio antenna.

In this case, the near-field electric and magnetic vector amplitudes are:

$$H\left[\frac{A}{m}\right] = \frac{Il}{4\pi D^2}$$

and

$$E\left[\frac{V}{m}\right] = \frac{Z_0 Il\lambda}{8\pi^2 D^3}$$

where l is the dipole length in meter

Contrasting with the near-field H a loop which falls with the inverse o D^3, the near-field H of a dipole falls c as $1/D^2$. Similarly, the near-field E of

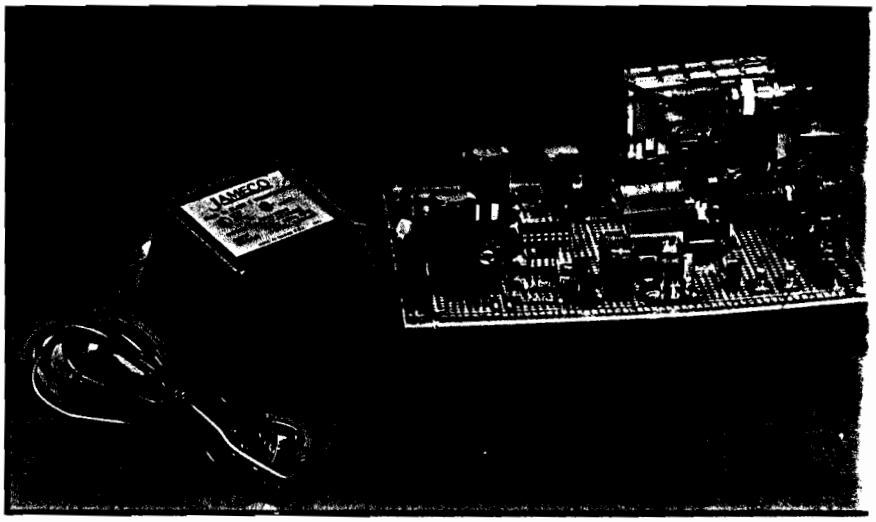


Photo 1—*A simple circuit can convert any triggered oscilloscope into a 100-kHz to 400-MHz spectrum analyzer suitable for near-field EMI sniffing.*

dipole falls off as $1/D^3$, in contrast to that of a loop, which falls as $1/D^2$.

The wave impedance of emissions radiated by a dipole is also affected differently by frequency:

$$Z_{wave} = \frac{Z_0 \lambda}{2\pi D} \quad [2]$$

Compare Equation 2 with Equation 1. The change in wave impedance as a function of frequency in the case of a dipole is inverse to that of a loop.

In the far field, the behavior of E- and H-fields is again similar to that of electromagnetic radiation from a loop. That is, they decrease as the observation distance increases:

$$H\left[\frac{A}{m}\right] = \frac{Il}{2\lambda D}$$

$$E\left[\frac{V}{m}\right] = \frac{Z_0 Il}{2\lambda D}$$

Beyond the transitional point, the wave impedance again remains constant at the value of Z_0. The result of a constant impedance in the far field means that the ratio of E to H components remains constant regardless of how the field generates.

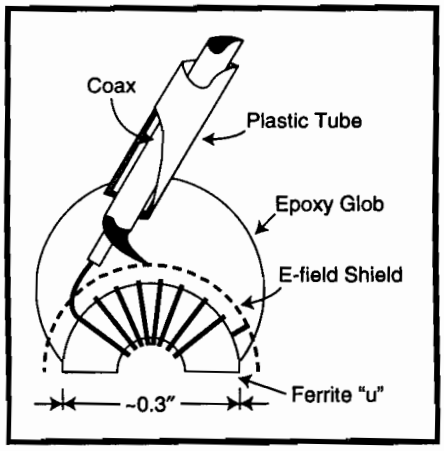


Figure 5—*A useful H-field probe can be constructed from a small, halved ferrite bead. Approximately 40 turns of thin enameled wire detect the magnetic flux concentrated by the ferrite. A small portion of the coax cable braid is used as an E-field shield for the coil. The assembly is mounted at the end of a small plastic tube which serves as a handle and is embedded in a glob of epoxy.*

Of course, real-life circuits are neither ideal open wires nor perfect loops, but hybrids of these two.

In a simplified form, as shown in Figure 3, a more realistic model of a circuit which radiates electromagnetic emissions assumes that an AC voltage source causes the flow of a current I in a rectangular loop enclosing an area S. The source impedance is Z_{source}, and the impedance of the load is Z_{load}, resulting in an overall equivalent impedance of $Z_{circuit} = Z_{source} + Z_{load}$.

In the near field, the electric- and magnetic-field vector magnitudes are given by:

$$E\left[\frac{V}{m}\right] = \frac{V_0 S}{4\pi D^3}$$

where $Z_{circuit}$ is greater than or equal to $7.9\ D[m]f[MHz]$, or

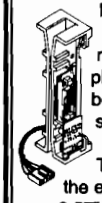
$$E\left[\frac{V}{m}\right] = \frac{0.631Sf[MHz]}{D^2}$$

where $Z_{circuit}$ is less than or equal to $7.9\ D[m]f[MHz]$, and

$$H\left[\frac{A}{m}\right] = \frac{IS}{4\pi D^3}$$

In the far-field, the electric- and magnetic-field vector magnitudes are

MAGNETIC CARD READER



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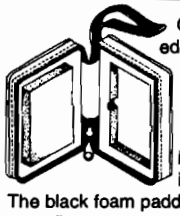

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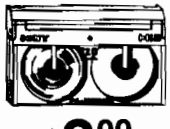

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given by:

$$H\left[\frac{A}{m}\right] = \frac{35 \times 10^{-6} IS\{f[MHz]\}^2}{D}$$

and

$$E\left[\frac{V}{m}\right] = Z_0 H = \frac{0.013 V_0 S\{f[MHz]\}^2}{D Z_{circuit}}$$

The second lesson of controlling radiated emission leaps out from these equations: keep the area enclosed by loops carrying strong time-varying currents to the minimum possible. Similarly, traces carrying high voltages should be kept as short as possible and be properly terminated.

Besides directing our attention to the parameters affecting radiated emissions, these equations are very useful when designing for compliance with EMI requirements.

As exemplified by Figure 4, near- and far-field ballpark estimates of EMI can be obtained from known circuit parameters for a large number of common circuit topologies.

PROBING E- AND H-FIELDS

As shown, the main reason why EMI standards require testing to be performed in the far field is that a constant impedance in the far field causes the ratio of *E* to *H* components to remain constant regardless of how the field was generated. Hence, measurements can be reproduced with reliability, and standardized methods of testing can be defined with ease.

From the past equations, however, it seems possible to establish a quantitative correlation to estimate far-field values from near-field measurements. Unfortunately, in practice, this is not the case.

Near-field measurements are extremely dependent on the source's exact geometry, the position of the near-field probe, and the interaction between the probe and the source. The variability of these values is too high to enable the exact measurements necessary to calculate radiation behavior in the far-field region.

Although it doesn't predict the outcome of compliance tests, near-field measurements are nevertheless useful for locating potential sources of radiated emissions. Here, near-field qualitative measurements with simple instruments can accurately pinpoint sources of EMI and identify their basic characteristics.

In essence, if a strong E-field and relatively weak H-field are detected from a certain circuit section, the culprit can usually be traced to a tr of high-voltage pulses on a long wi an unterminated line, or a trace dri ing a high-impedance load.

Conversely, if the H-field is stro and the E-field probe is inactive, th source of EMI is most likely a loop like circuit where strong currents circulate. Some examples are PCB tracks carrying strong currents, ind tors in switching power supplies, a eddy currents induced in metal en sures by strong internal fields.

Since the same equations descri radiation emission apply to the rec tion of emissions, it is apparent th small loop of wire can act as a nea field probe mostly sensitive to H-fields.

In contrast, E-fields are detecte preferably by a short exposed wire Measurements are then taken witl wideband AC voltmeter or a spect analyzer.

Even a simple single-turn wire l at the end of a coax cable can be a effective H-field probe. With this a rangement, maximum output fron the probe is recorded when the loo

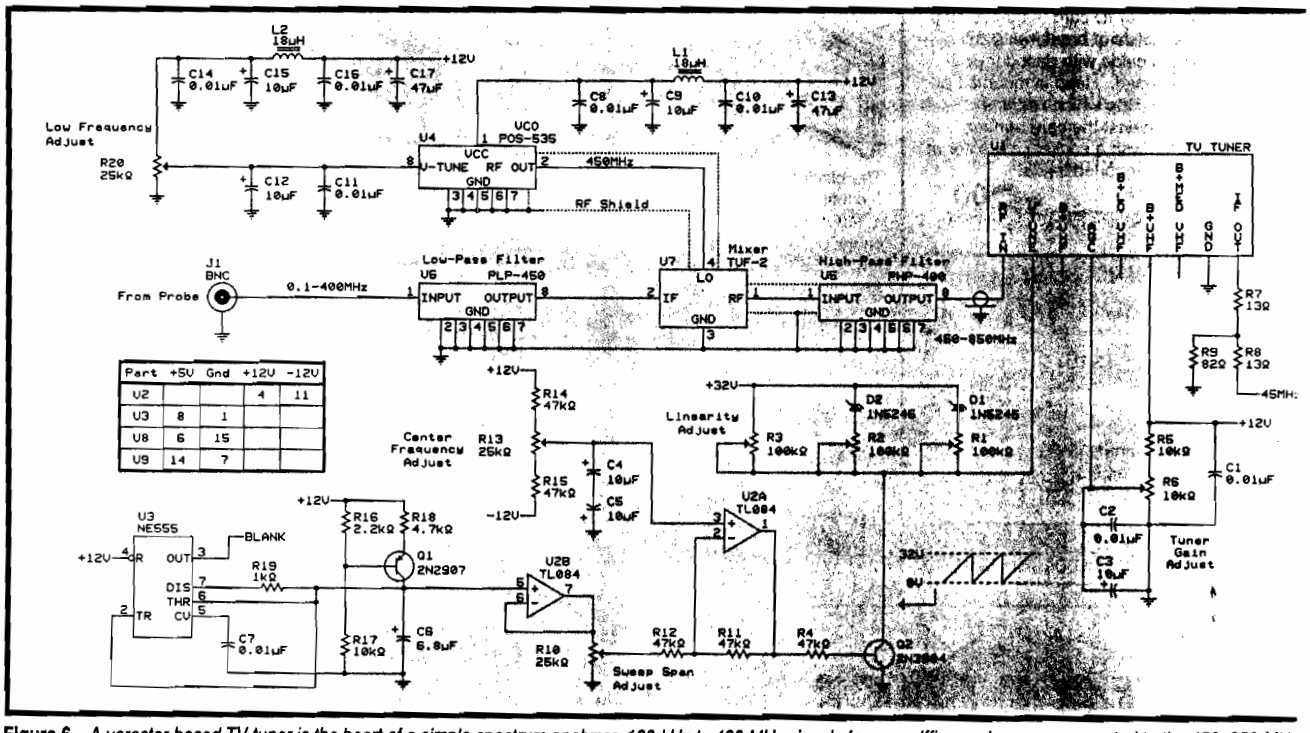

Figure 6—*A varactor-based TV tuner is the heart of a simple spectrum analyzer. 100-kHz to 400-MHz signals from a sniffing probe are upconverted to the 450–850-MHz band, where the tuner can be swept by a sawtooth waveform. The tuner produces a 45-MHz intermediate frequency which can be processed to derive the input signal st Direct connection of the probe to the tuner input extends the range of the spectrum analyzer to the high-VHF/UHF region (450–850 MHz).*

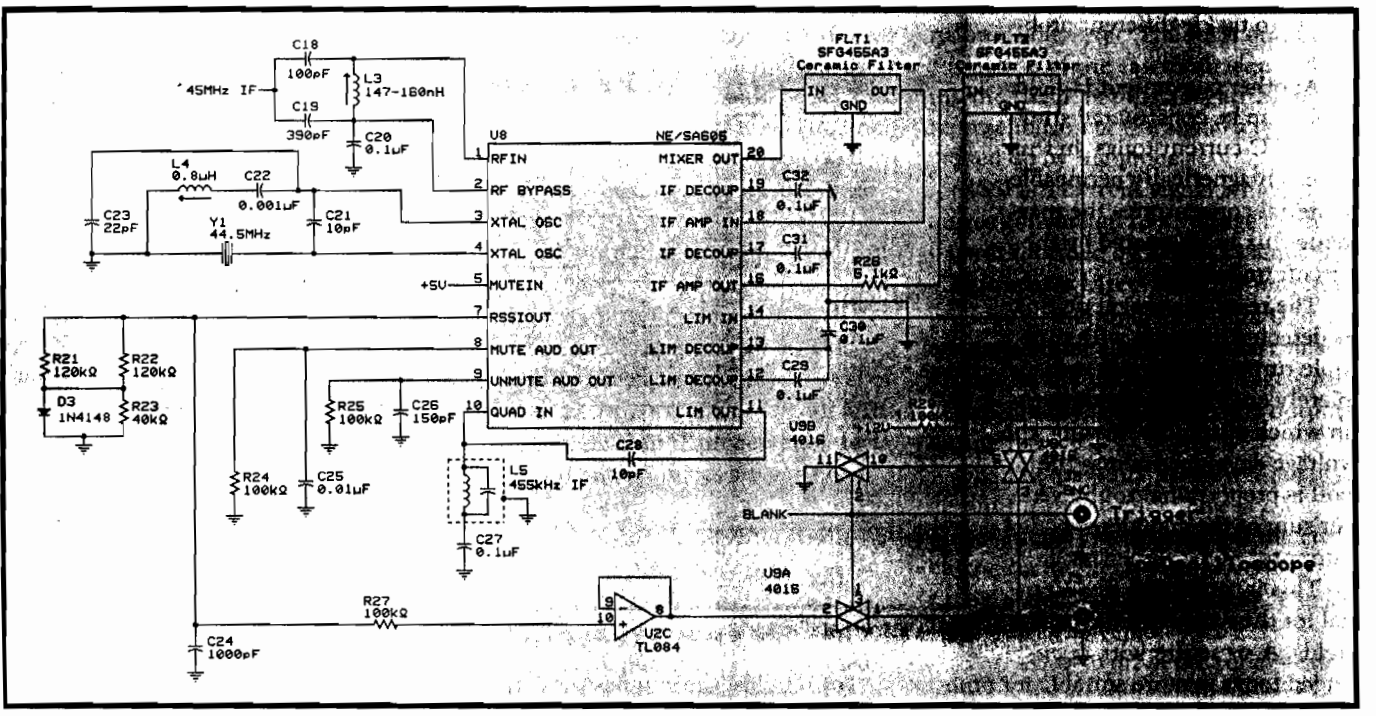

Figure 7—*The intermediate frequency output of the tuner is detected by U8, a single-chip IF processor. The Received Signal Strength Indicator (RSSI) output, as a function of the sawtooth signal driving the tuner, is a logarithmic representation of the spectrum of the signal picked up by the probe.*

in immediate proximity and aligned with a current-carrying wire. This directionality is very useful for pinpointing the exact source of a suspicious signal.

The diameter of the loop makes a large difference on H-field measurements [4]. The area enclosed by the loop influences the sensitivity of the probe since it determines the number of magnetic-flux lines which are intercepted to produce a detectable signal.

A larger loop obviously develops a larger voltage at the input of the voltmeter or spectrum analyzer. On the other hand, larger loops have inherently larger self-inductance and equivalent capacitance than smaller loops.

As inductance increases, the network formed with the complex impedance of the measurement setup resonates at lower frequencies, beyond which the probe cannot be used. Moreover, larger loops make it much more difficult to identify the exact source of an interfering signal. Their size does not allow them to selectively pick up radiations from single lines when a multitude of the latter are clustered close together.

Coils with multiple turns increase sensitivity without appreciably increasing the physical size of the coil. However, this solution results in reduced spectral response due to increased self-inductance.

Loop geometry must therefore be chosen each time by compromise. It's a good idea to keep a variety of probes

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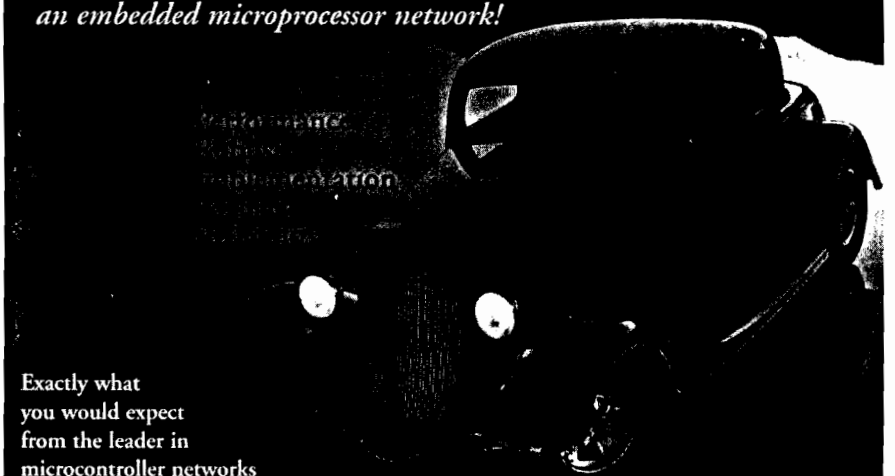

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handy to tackle different problems.

Another convenient H-field probe can be constructed similar to AC-current tongs. In this case, a magnetically permeable material concentrates the magnetic-flux lines created by the circuit under test.

The resulting magnetic flux is detected by a coil with multiple turns. If the tongs completely enclose the conductor in which current is flowing, the voltage developed across the coil is proportional to the vector sum of the conductor's currents.

This is, of course, impractical for the needs of sniffing H-fields. A structure with open-ended tongs is more suitable for probing a circuit without modifying it.

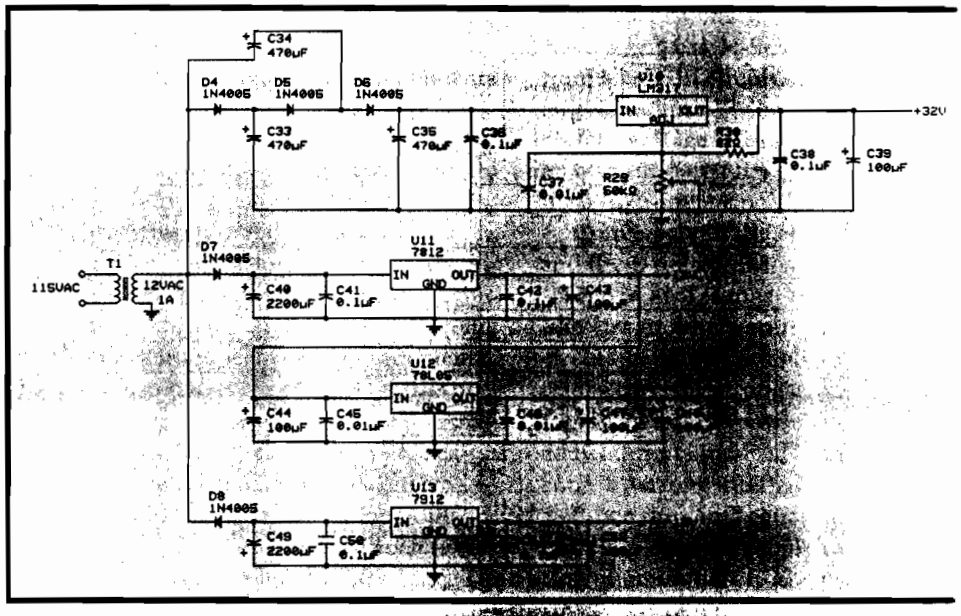

Figure 8—*DC power for the various circuits of the spectrum analyzer is derived from a single 12-VAC input.*

The probe can be built as shown in Figure 5, using a small ferrite bead (e.g., 0.1" thick, 0.3" outer diameter) cut in half. The construction depends on the ferrite selected, but generally, 40–50 turns of thin enameled copper wire provides suitable sensitivity.

Solder the coil terminals to the center and shield of a coax cable. After insulating the central conductor connection, use a portion of the braid to cover the assembly. This shields the coil from the E-field.

The assembly can then be mounted at the end of a small plastic tube and embedded within a glob of epoxy. For the prototype probe, I measured a virtually flat bandwidth from approximately 600 kHz to 10 MHz.

Better bandwidth is achieved with a VCR magnetic head. Video heads detect broadband magnetic fluctuations. For this reason, they are useful for sniffing H-fields from 2 to 120 MHz with relatively flat response.

To construct the probe, carefully remove one of the magnetic heads from a discarded drum. Even a worn-out head works well.

Soiled heads should be cleaned with a swab and pure alcohol. Degaussing also improves an old head's sensitivity. All other aspects of constructing and using this probe are the same as for the ferrite-bead probe.

For E-fields, the simplest near-field probe is a coax cable in which a short segment of the center conductor extends beyond the braid at the unterminated end of the coax. Similar to the loop probe, a longer wire picks up a stronger signal at the expense of specificity and bandwidth.

In general, select a wire length which gives a sensitivity of approximately 3 mV/m. At this level, potentially problematic emissions can be identified without causing undue concern about low-level emissions.

Constructing the ideal H- or E-field probe for a specific job takes some trial and error, since the effort of electromagnetic modeling required for proper design is overkill in most applications.

You may nevertheless want to check whether a probe resonates within the desired spectral range. Connect a wideband probe to an RF generator set to track the tuning frequency of a wideband spectrum analyzer. The probe should be located in close proximity to the emitting probe, and connected to the spectrum analyzer's input. The limit of the probe's useful bandwidth is the point at which the first abrupt resonance appears.

Before plugging the spectrum analyzer into the powerline, however, the first step in conducting a near-field EMI study is to draw the assembly's component placement diagram. The diagram should indicate circuit points identified in the mathematical circui harmonic analysis as potential sourc for EMI radiation. Only after this pre liminary work should you try bench testing.

Conduct a coarse near-field sweep at relatively high gain to identify EM hot spots. Log the frequencies at whi strong components appear when scan ning the unit under test. Detailed scanning, using a more discriminatin probe, can then concentrate on the h spots identifying the culprit circuit generating offending emissions.

You gain a valuable source of clue for future troubleshooting by printin the spectral estimate at each point th measurements highly agree or strong disagree with the circuit's harmonic analysis.

In any case, keep detailed and orga nized notes of the near-field scans. They prove invaluable when attempt ing quick fixes while the clock is run ning at the far-field compliance-testi facility.

A BARE-BONES SPECTRUM ANALYZER

While an AC voltmeter indicates the field strength a probe is exposed t it doesn't indicate an emission's spec tral content. A spectrum analyzer is a tool that certainly can't be beat in th search for offending signals.

Unfortunately, spectrum analyzer are often beyond tight budgets. For

near-field sniffing, however, even the crudest spectrum analyzer does a magnificent job.

Photo 1 shows a simple home-brewed adapter that converts any triggered oscilloscope into a spectrum analyzer that gives qualitative spectral estimates of 100 kHz to 400 MHz.

As shown in Figure 6, a voltage-controlled TV tuner U1 forms the basis of the simple spectrum analyzer. Most any voltage-controlled tuner works. You may be able to get one free from a discarded TV or VCR PCB. tion vary from device to device, but the pinout is usually identified by stampings on the metallic can of the device.

Varactor-controlled TV tuners receive signals on their RF input at a frequency determined by the voltage applied to the VTUNE input. With power applied to the UHF section of a tuner, typical control voltages between 0 and 32 V span a frequency range of approximately 450–850 MHz.

Tuner sensitivity can be adjusted through the AGC input. The output of the tuner is a standard 45-MHz IF.

However, the 450–850-MHz range is not directly applicable to the large bulk of EMI sniffing. For this reason, a more appropriate range of 100 kHz to 400 MHz is converted up to the tuner's input range through a circuit formed by U4–U7.

Here, signals from the probe are low-pass filtered by U6 and injected into the IF port of a TUF-2 mixer. The LO input of the mixer is fed with the output of U4, a self-contained voltage-controlled oscillator tuned to 450 MHz by potentiometer R20.

The RF port of the mixer outputs signals with frequency components at the sum and difference between the IF input and the LO frequency. This output is high-pass filtered by U5 to ensure that only up-converted components are fed to the tuner input.

Sweeping the tuner across its range is accomplished by a sawtooth waveform which spans approximately 1–31 V. The basic sawtooth is generated by U3 and Q1, and buffered by U2b. The span of the sawtooth is set by attenuator R10, while the center of the sweep is adjusted by introducing an offset on U2a by means of R13.

The output of U2a is amplified by transistor Q2, which should be selected for a gain of 50 or less. The final span and linearity of the sweep is adjusted in three ranges by R1, R2, and R3.

The IF output of the tuner is attenuated to a level suitable for processing by the circuit in Figure 7. Select the actual value of the resistors for this attenuator based on the output level of the specific tuner you use. U8, an NE/ the signal and produces a logarithmic output of signal strength.

In this portion of the circuit, the 45-MHz IF signal is coupled to the input of a RF mixer internal to U8 via a tuned circuit formed by C18, C19, and L3. The LO input of this mixer is fed from a 44.5-MHz crystal-controlled oscillator. The resulting 455-kHz IF is filtered by two ceramic filters, FLT1 and FLT2.

An internal Received Signal Strength Indicator (RSSI) circuit is used as a detector and linear-to-logarithmic converter. The RSSI output, as a function of the sawtooth signal driving the tuner, is thus a logarithmic representation of the signal's spectrum picked up by the probe.

RSSI is a current signal which requires conversion to a voltage by the network formed by R21–R23 and D3. C24 low-pass filters the RSSI output to produce a smooth display, and U2c acts as a buffer and impedance transformer for the current-to-voltage converter. Finally, U9a blanks the output

Figure 8 presents the power-supply circuit for the adapter. The ±12-V power supply powers most of the circuitry, including the upconverter, tuner, and sawtooth generator. The IF processor is powered by +5 V.

The +32 V to drive the tuner's varactors is obtained by stepping up the 12-VAC input to +48 V and then reaching the desired voltage through U10, an LM317-adjustable linear regulator.

To operate the spectrum analyzer, the Y output of the adapter is con-

#115

nected to the vertical input of the oscilloscope, and the TRIGGER output is connected to the trigger synchronization input of the scope.

The horizontal frequency of the oscilloscope is set so one full sweep caused by the sawtooth fits the full graticule on the oscilloscope's screen. Fine-tune it by either trimming the time base of the scope or by appropriately adjusting the value of R18.

Alternatively, a two-channel oscilloscope can be operated in the X–Y mode by injecting the sawtooth available at pin 7 of U2a to the appropriately scaled *x*-axis channel.

The comb generator circuit in Figure 9 calibrates the adapter. The circuit is simply a TTL-compatible 40-MHz crystal-controlled oscillator module feeding a synchronous binary counter.

It is called a comb generator because the spectral pattern of any of its outputs resembles a hair comb with its prongs pointing up. Because these spectral components occur at harmonic multiples of the selected fundamental square-wave frequency, it follows that the frequency difference between consecutive prongs must be the same as the value of the fundamental frequency of the square wave.

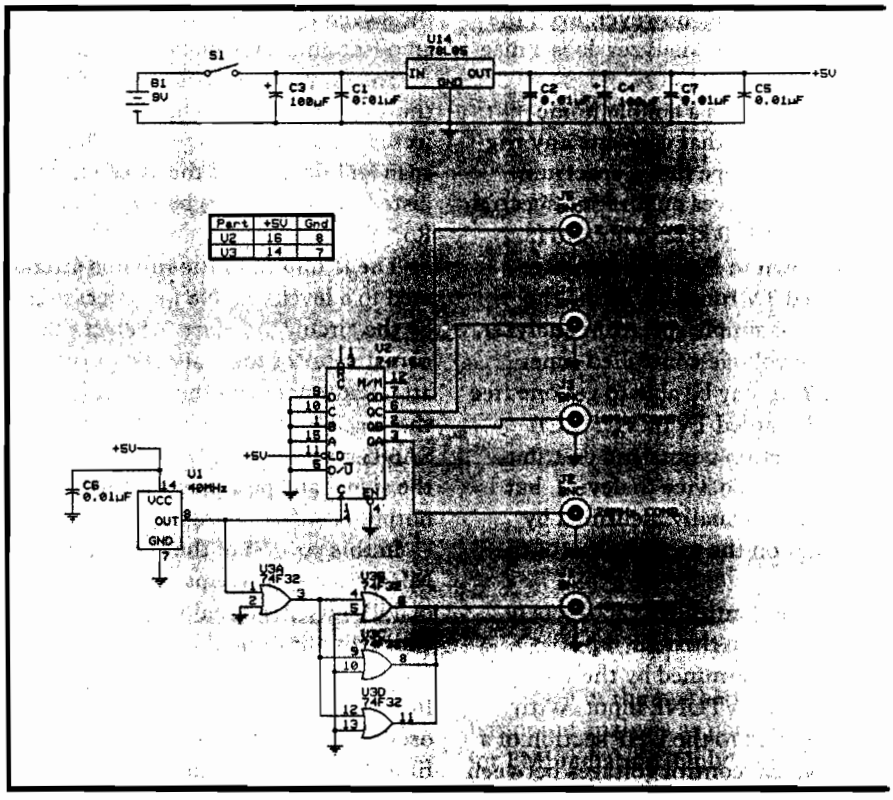

Figure 9—*High-frequency clocks and fast logic generate broadband signals extending well into the hundreds-of-megahertz region. This generator produces various comb patterns which help calibrate spectrum analyzers.*

The graph in Figure 10 presents the pattern obtained when the 20-MHz comb output of the generator is probed by a commercial-grade spectrum analyzer.

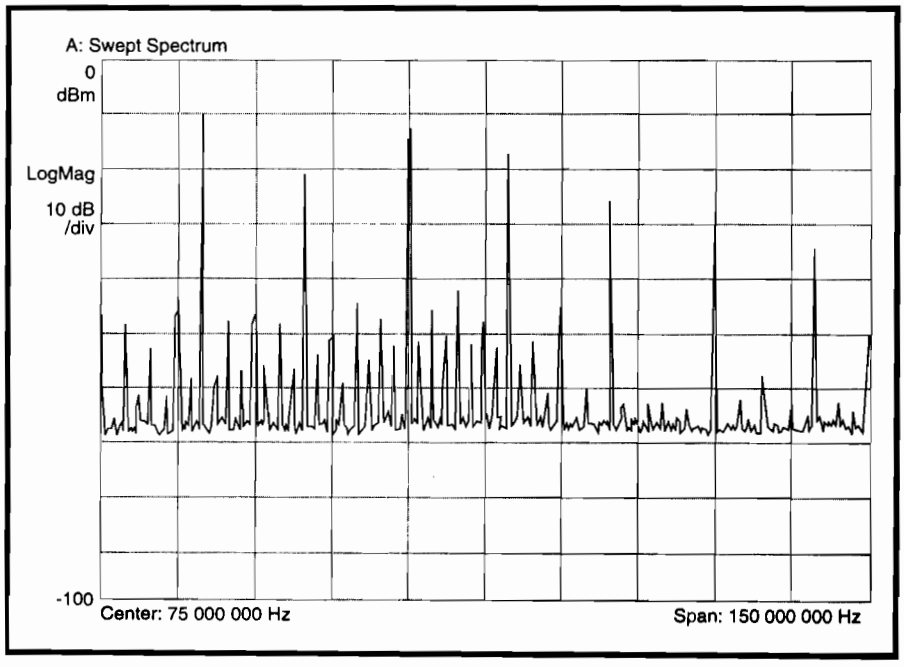

Figure 10—*The spectral pattern obtained from the output of the comb generator serves as a frequency ruler because it presents strong spectral lines at every harmonic of the fundamental square wave. Notice the similarity between the envelope formed by the spectral components of this 20-MHz comb and the nomograms of Figures 1 and 2.*

I accomplished AC coupling through a series-connection 15-pF capacitor. The capacitor's output side is terminated to ground through a 50-Ω noninductive resistor. This is the gold standard against which the adapter should be calibrated.

Start testing the adapter by setting the sawtooth generator to vary the voltage at the VTUNE input of the tuner between 1 and 31 V. Initially, set R6 to apply 2.5 VDC to the AGC pin of U1.

Adjust the upconverter LO frequency to 450 MHz by trimming R20. At U4's VTUNE input, 9.6 VDC typically results in the desired LO frequency. L4 should be trimmed to achieve stable oscillation of the 44.5-MHz IF LO oscillator.

With a 40-MHz comb applied to the input of the adapter through a 15-pF coupling capacitor and with 50-Ω termination, adjust L3 to approximate the expected 40-MHz comb pattern on the oscilloscope.

After achieving a satisfactory display for the 40-MHz comb, calibrate the linearity of the adapter using a 20-MHz comb by first trimming R3 to

produce equal spacing between spectral lines throughout the lower third of the display. Then, linearize the mid-range by trimming R2, and finally the high range by trimming R1.

PASSING THE TEST

Designing equipment that passes EMI and EMC compliance testing without fixes or delays never happens by mistake. Rather, it involves considering compliance with EMI and EMC regulations from the very beginning of product formulation and design.

Developing a first prototype free of foreseeable trouble gives a head start in the battle against EMI. Such development is possible by carefully selecting the technologies which fulfill the product requirements while minimizing EMI, using good design and construction practices, and making extensive use of circuit simulation tools.

Near-field probing of the first prototype should reveal real-world EMI effects that escape from the limited view of initial modeling. Correcting any problems through filtering, shielding, or redesign is inexpensive at early design stages. The second prototype already has a good chance of passing compliance testing with minimal rework.

This article presented only a few of the ways in which technology selection, circuit design, and layout techniques influence the generation of EMI. Many more books [5,6,7] and articles [8] disclose the secrets of the EMI and EMC world, all the way from Maxwell's equations, through the legalities of regulation, and into the tricks of the trade for taming EMI.

Considering the stiff economical, technical, and legal penalties brought by manufacturing a product that does not comply with EMI and EMC regulations, you should be motivated to keep EMI in sight at every turn of the design process.

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REFERENCES

[1] G. Dash and I. Strauss, "Inside Part 15: Digital Device Approval," *Compliance Eng.*, 1995 Annual Reference Guide, A11–A18.

[2] R.W. Gubisch, "The European Union's EMC Directive," *Compliance Eng.*, 1995 Annual Reference Guide, A55–A64.

[3] G. Dash and I. Strauss, "How the FCC Enforces Part 15," *Compliance Eng.*, 1995 Annual Reference Guide, A21–A27.

[4] V. Kraz, "Near-Field Methods of Locating EMI Sources," *Compliance Eng.*, 43–51, May/June 1995.

[5] M. Mardiguian, *Controlling Radiated Emissions by Design*, Van Nostran Reinhold, New York, NY, 1992.

[6] M. Mardiguian, *Interference Control in Computers and MicroprocessorBased Equipment*, Don White Consultants, Gainesville, VA, 1989.

[7] C. Marshman, *The guide to the EMC Directive*, IEEE Press, Piscataway, NJ, 1992.

[8] I. Strauss and G. Dash, Various Articles on "Design for Compliance," *Compliance Eng.*, 1995 Annual Reference Guide, A115–A154.

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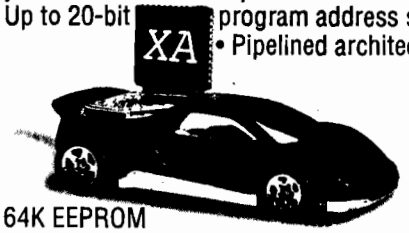

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#116